New frequency counting principle improves resolution

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Abstract: Frequency counters have gone through several evolution stages in their design since their first appearance on the market:

Stage 1	until 70-ies	Conventional counting
Stage 2	1980-ies	Reciprocal counting (period measurement + inversion)
Stage 3	1990-ies	Interpolating Reciprocal Counting
Stage 4	2000-ies	Multiple Time Stamp Average Continuous Counting

This paper describes the theory and design of frequency counters, and analyses the improvements in the latest generation of frequency counters.

The newly introduced high-resolution CNT-90 Timer/Counter/Analyzer from Pendulum Instruments AB in Sweden is used as commercially available example of the latest design technology in this presentation.

The advantages of continuous time-stamping technique are discussed, for example regression analysis to reduce effects of measurement noise and to improve resolution to 12 digits for 1s of measuring time. Another example is the ability for seamless back-to-back measurements without missing any period, which is essential for theoretically correct calculation of Allan Deviation.
Keywords: Time-stamping, linear regression

1. Introduction

An ideal sine wave signal (carrier wave) $U(t) = A \cdot \sin(2\pi ft)$ has a constant frequency vs time behaviour, $f = f_0 = constant$. This means that the signal phase grows linearly with time:

$$\Phi(t) = 2\pi f_0 t + \Phi_0 \qquad [1.1]$$

The *mean* frequency of a continuous periodic signal over a certain *measurement time*, is illustrated in figure 1.

$$frequency\ (Hz) = \frac{number\ of\ complete\ cycles}{measurement\ time} \qquad [1.2]$$

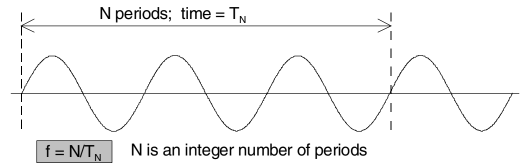

Figure 1. Definition of mean frequency

Even if the signal were assumed to be ideal, in the real world, it would be subject to noise processes and interference, which means that individual periods could vary. The concept of mean frequency involves integral numbers of counted signal cycles, at least one cycle. You cannot define a signals frequency by measuring a fraction of a cycle.

Real world signals do not have constant stable frequencies, not even clock oscillators as will be discussed in section 7. There could be modulated signals, frequency hopping signals, swept frequency signals, burst signals and much more. Then f is a function of time $f(t)$ and the phase function in [1.1] is expressed differently.

$$\Phi(t) = 2\pi f(t)t + \Phi_0 \qquad [1.3]$$

Please note that the concept of mean frequency may be useless for these types of signals. The average frequency over 80-channels WLAN using FHSS or the average of several burst cycles containing chirp radar frequency is not meaningful. Instead the challenge is to closely follow and represent the actual frequency over time $f(t)$ inside the burst or alternatively the statistical distribution of WLAN channels. This requires very fast and high-resolution measurements, found in very few frequency measurement devices today, one being the new CNT-90 Timer/Counter/Analyzer (Pendulum Instruments AB).

The rest of this paper will focus on frequency measurements on stable signals, where the concept of mean frequency is meaningful.

2. Conventional counters

Conventional counting was the first frequency counting method and these counters did not measure according to the definition of frequency above [1.2]. The conventional principle is to open an exact 1-second-gate and count the number of input cycle trigger events that occur during that second. The counting register contains the number of cycles counted during exactly one second, which is a sort of frequency (cycles/s). The precisely defined 1s gate-time is derived from a X-tal oscillator reference (usually a 10 MHz signal) with a good accuracy.

Gate time is *not* synchronized with the input signal. The uncertainty of the measurement is ±1 input cycle count, which means that the resolution is 1 Hz during a 1s gate time for *all* input signal frequencies. To allow measurements with a resolution other than 1 Hz, gate-times of a multiple (or a sub-multiple) of 1s are used. E.g. a gate time of 10 s will increase resolution tenfold and add one more digit to the read-out.

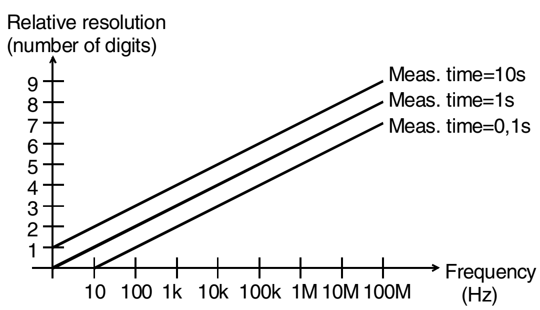

Figure 2. Resolution of a conventional counter is bad for low-medium frequencies and adequate for high frequencies only

3. Reciprocal counting

In the early 1980-ies, microcomputer based instruments started to use reciprocal counting. The input signal trigger, and not the internal oscillator, controls the gating of a multi-period average measurement. N input signal periods are counted during measurement time MT. They calculate mean cycle time $\overline{T} = MT/N$ and the *reciprocal* value; mean frequency $\bar{f} = 1/\overline{T}$.

Figure 3 shows the block diagram of a first generation reciprocal frequency counter. It contains two counting registers. One counts the number of input cycles and the other counts the clock pulses, to measure the time duration. Two synchronized main gates simultaneously control both counting registers.

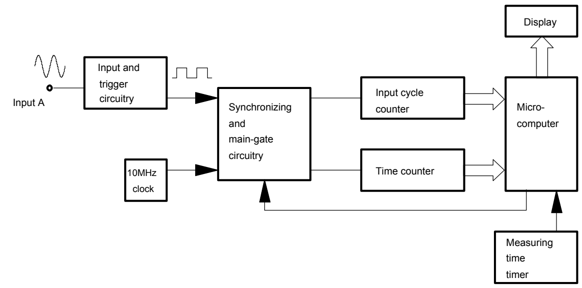

Figure 3. Block diagram for a reciprocal frequency counter

Unlike conventional frequency counters, the set measuring time is not an exactly defined gate time. The *desired* measuring time is set via the microcomputer, but the *actual* measuring time MT is synchronized to the input signal triggering. The measurement contains an exact number of input cycles. Thus the ±1 input cycle error is avoided. Truncation errors are now in the time count; i.e. ±1 clock pulse.

To obtain the mean frequency value, the following division is made:

$$frequency = \frac{Counted\ input\ cycles}{(Counted\ clock\ pulses) \times t_c} = \frac{N}{MT}$$

Where t_c is the time of one clock cycle

The relative resolution of the calculated result is:
$resolution = \pm\, t_c / MT$ normally $\pm 100\ \text{ns}/MT$.

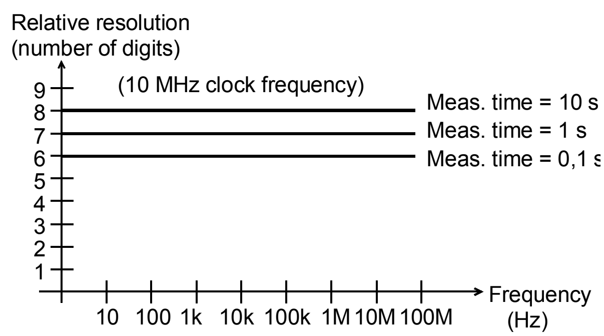

Figure 4. The relative resolution is independent of input frequency for a reciprocal frequency counter

To obtain a higher resolution, one could increase the clock frequency. E.g. a 100 MHz reference clock would give a relative resolution of $\pm 10\,\text{ns}/MT$, and thus one digit more in the displayed result, compared to a standard 10 MHz reference clock.

4. Interpolating reciprocal counting

The resolution of the 2:nd generation reciprocal counters is always $\frac{\pm 1 \text{ clock period}}{\text{Measurement time}}$.

In the third generation of counters, resolution is improved by means of *analog interpolation* of the fractional clock pulse. Instead of just counting the clock pulse edges to determine the time between start and stop trigger, also the *fractional* clock pulse in the beginning and end of the measurement is captured.

Figure 5 shows the block diagram of an interpolating frequency counter, like the Pendulum CNT-85. Compared to the basic reciprocal counter (fig. 3) such a counter contain also two interpolators, one for the start trigger event and one for the stop event.

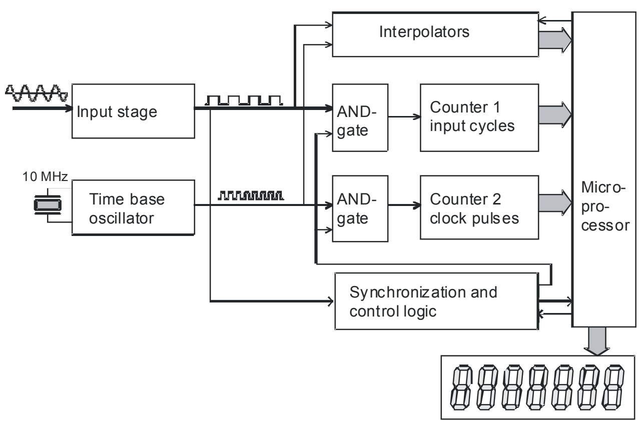

Figure 5. Block diagram for an interpolating reciprocal frequency counter

Figure 6 illustrates the interpolator's principle to capture the short fractional time between the start trigger and the following clock pulses, respectively the stop trigger and the following clock pulses.

The analog interpolator in figure 6 starts to charge the capacitor, with a constant current I, at the arrival of the trigger event and stops on the 2:nd following clock pulse. The capacitor is charged as $Q(t) = I \cdot t$. The voltage (U) over the capacitor is:

$$U(t) = Q(t)/C = (I/C) \cdot t \quad [4.1]$$

The charge time (t) varies between 1 and 2 clock cycles, normally 100 to 200 ns. $U(t)$ also varies between U_0 (charge time is 1 clock cycle) and $2U_0$ (2 clock cycles). By selecting I and C ($U_0 = (I/C) \cdot t_c$), you can reach a convenient range (some Volts). The interpolator circuitry is duplicated for the stop trigger.

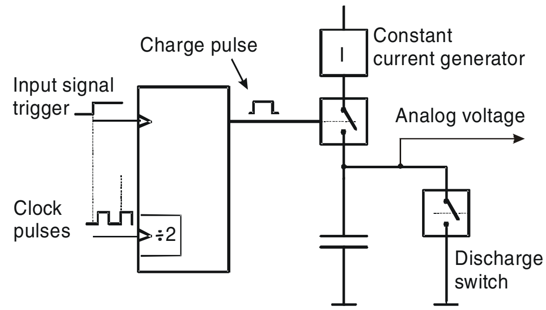

Figure 6. The basic principle of an analog interpolator is a time-to-voltage-conversion

The number of cycles counted N is as before an exact integer number, but the accuracy of the corresponding time (MT) is significantly improved. Instead of a resolution of ±1 clock cycle, as in classical reciprocal counting, the interpolated resolution is improved to less than a percent of a clock cycle. MT is calculated as $T_N + T_1 - T_2$, where T_N is the digitally counted time (number of clock cycles), T_1 is the interpolated fractional clock pulse between start trigger and following clock pulse and T_2 is the fractional clock pulse between stop trigger and following clock pulse.

The advent of interpolating reciprocal counters typically improved the time interval, or single period, resolution with 100 to 400 times, from 100 ns (single-shot) to 1 ns and below for a timer/counter with a 10 MHz time base oscillator. The Pendulum timer/counter model CNT-81 combines interpolation techniques with 10 times increased clock frequency (100 MHz vs. 10 MHz), and reaches 50 ps resolution as single period or time-interval resolution. This corresponds to a relative resolution in frequency measurements of $50\,\text{ps}/MT$ (rms value), approx 1000 times improvement compared to typical 2:nd generation reciprocal counters ±100 ns/MT (limit value).

5. Continuous time stamping and statistical improvements

In reciprocal counters, with or without interpolation techniques, a frequency measurement has a defined start (= start trigger event), and a stop (= stop trigger event) plus a dead-time between measurements to read out and clear registers, do interpolation measurements and prepare for next measurement. *Continuous time stamping* changed that scenario.

In a *time-stamping counter*, the input trigger events, and the clock cycles, are continuously counted, without being reset. At regular intervals, pacing intervals, the momentary contents of the event count register and time count register is transferred to the memory. The read-out of register contents is always synchronized to the input trigger, so it is the trigger event that is time stamped. Each stored time stamp is also interpolated "on the fly" for improved resolution. The contents in the memory is thereafter post-processed.

A one-second frequency measurement in a fast processing counter could contain hundreds or thousands of paced time-stamped events, not just a start event plus a stop event. This makes it possible to use *linear regression using the least-squares line fitting* to further improve accuracy. See figure 7.

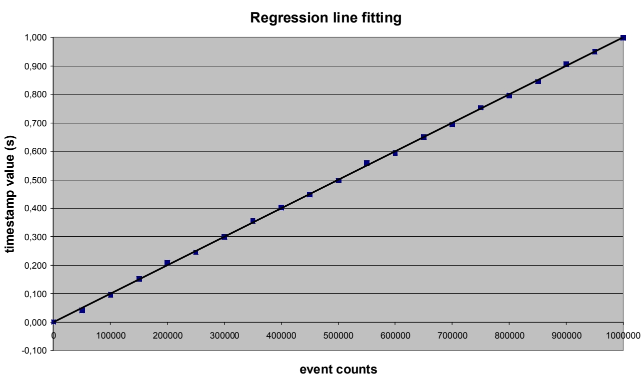

Figure 7. Time-stamping counters have a lot of intermediate time stamps of trigger events between the start and the stop of measurement

We have a series of data $\{x_k, y_k\}$, where x_k is the accumulated contents of the Event count register and y_k is the accumulated time at each sample point. The estimated frequency f^* is the inverse slope of the line that best fits this data set. Each x_k is an exact number, whereas each y_k has a basic uncertainty t_{RES}.

Our problem is to find the best estimate of the mean frequency over measurement time *MT*, by finding the straight line $y = a + b \cdot (x - x_0)$ using linear regression, where:

x = number of cycles counted (independent variable)
y = elapsed accumulated time (dependent variable)
a = time value for $x = x_0$ (first sample)
$b = f^{*-1} = T^*$, the slope of the regression line is the estimated mean period T^* or the inverse value of the estimated mean frequency f^{*-1}. From now on we will substitute the slope b with T^*.

From basic statistics we know that the regression line slope b (T^*) is calculated as:

$$T^* = \frac{n\sum x_k y_k - \sum x_k \sum y_k}{n\sum x_k^2 - \left(\sum x_k\right)^2} \qquad [5.1]$$

and that the variance of the slope b (or T^*) is

$$s^2(T^*) = \frac{s^2(y)}{\sum (x_i - \bar{x})^2} = \frac{s^2(y)}{s^2(x) \cdot (n-2)} \qquad [5.2]$$

$s(y)$ is the normal rms-resolution t_{RES} for a single time stamp, but what is $s(x)$? The independent variable X is assumed to have a linearly increasing distribution over the range x_0 to $x_0 + N$, with samples that are evenly spread over the full interval, that is

$$x_k = x_0 + \frac{kN}{n} \qquad [5.3]$$

For large values of n (number of samples), we can approximate the distribution with the continuous rectangular distribution with a density (probability) function of:

$$p(x) = \begin{cases} 0 & x < x_0,\ x > x_0 + N \\ 1/N & x_0 \le x \le x_0 + N \end{cases} \qquad [5.4]$$

For such a distribution we find

$$\mu = \int_{-\infty}^{+\infty} x p(x) dx = \frac{1}{N}\int_{x0}^{x0+N} x dx = x_0 + \frac{N}{2} \qquad [5.5]$$

$$\sigma^2 = \int_{-\infty}^{+\infty} (x-\mu)^2 p(x) dx = \frac{1}{N}\int_{x0}^{x0+N} (x - x_0 - \frac{N}{2})^2 dx = \frac{N^2}{12} \qquad [5.6]$$

Thus the standard deviation $s(x)$ for the discrete variable $X = \{x_k\}_0^n$ can be approximated to

$s(x) \approx \sigma = \frac{N}{2\sqrt{3}}$ for large values of n

This approximation, plus the knowledge that $T^* = MT / N$ and $s(y) = t_{RES}$, gives us the variance of the slope of the regression line from [5.2]:

$$s^2(T^*) = \frac{s^2(y)}{s^2(x) \cdot (n-2)} = \frac{1}{N^2} \cdot \frac{12 s^2(y)}{n-2} = \frac{T^{*2}}{MT^2} \cdot \frac{12 t_{RES}^2}{n-2} \qquad [5.7]$$

which finally leads us to the *relative* period or frequency uncertainty:

$$\frac{s(T^*)}{T^*} = \frac{s(f^*)}{f^*} = \frac{2\sqrt{3} \cdot t_{RES}}{MT \cdot \sqrt{n-2}} \qquad [5.8]$$

Using linear regression analysis (least square line fitting), gives a better estimate than just using the two end points for calculation and improves the relative resolution of the estimated frequency ($\Delta f/f^*$) from $\frac{\sqrt{2}\cdot t_{RES}}{MT}$ to $\frac{2\cdot\sqrt{3}\cdot t_{RES}}{MT\cdot\sqrt{n-2}}$, where:

t_{RES} = individual timestamp uncertainty

MT = Measuring Time

n = Number of event/timestamp value pairs used in the calculation.

The improvement in resolution between the two methods is thus:

$$\frac{2\cdot\sqrt{3}}{\sqrt{2}\cdot\sqrt{n-2}}=\frac{\sqrt{6}}{\sqrt{n-2}}\approx\frac{2.45}{\sqrt{n}}, \text{ for n>>2.} \qquad [5.9]$$

What if n is small, lets say n = 6? Then the approximation of a continuous rectangular distribution is no longer correct, and the standard deviation needs to be calculated based on discrete samples, which can be shown to give better resolution improvement than the approximation [5.9], for small values of n.

Example:

In the Pendulum model CNT-90, that uses the linear regression resolution improvement method, the random uncertainty is:

$$\frac{2\sqrt{3(t_{RES}^2+(trigger\ error)^2)}}{MT\cdot\sqrt{n-2}}\times Frequency\ or\ Period \qquad [5.10]$$

t_{RES} = 70 ps and $n=\frac{800}{MT}$.

Trigger error is the effect of superimposed noise on the input signal, which can be neglected for ideal square wave signals. If we assume no contribution from trigger errors and a MT of 1s (n=800), we get a relative resolution of:

$$\frac{2\sqrt{3}\cdot 7\cdot 10^{-11}}{1\cdot\sqrt{798}}\approx 8.6\cdot 10^{-12}$$

The CNT-90 can also use traditional frequency calculation (using start/stop only), with the random uncertainty of:

$$\frac{\sqrt{2(t_{RES}^2+(trigger\ error)^2)}}{MT}\times Frequency\ or\ Period \qquad [5.11]$$

which would give for MT = 1s, a relative uncertainty of $\sqrt{2}\cdot 70\text{ps}/1\text{s}\approx 1\cdot 10^{-10}$

The random uncertainty in this example is affected as predicted in [5.9] with the factor of $2.45/\sqrt{n-2}\approx 0.086$. Resolution is thus improved from 1E-10 (start-stop) to 8.6E-12 (regression).

The CNT-90 counter has an automatic mode, where the regression line fitting is executed at measuring times ≥200 ms and the number of samples used in the calculation are gradually reduced as measuring time increases $\left(n=\frac{800}{MT}\right)$. This gives the following resolution curve for CNT-90, see figure 8, where the dashed line is the traditional start-stop method. The resolution is improved for measurement times MT up to approx 100s and at 1s measuring time the improved resolution is typically 6E-12.

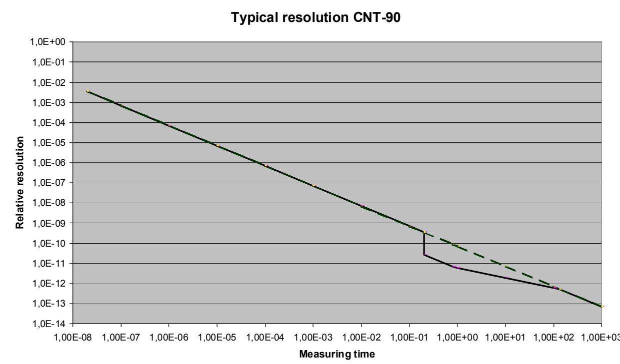

Figure 8. Resolution of the continuously time stamping counter CNT-90. Resolution improvement via linear regression occurs from measuring times from 200 ms up to 100s (auto mode)

6. Comparison of linear regression and traditional start-stop methods

One might believe that linear regression always is superior to traditional start-stop type of frequency measurements. But there are some limitations.

One obvious draw-back is that this post-processing of hundreds of sample data is time consuming, even if the raw data collection is fast. That means that the measuring speed of the frequency counter is reduced.

Linear regression is very useful, to reduce *random* noise in the measurement process, whether this noise is internally generated in the measuring device or externally added to the measurement signal. This method also assumes that the best fit is a *linear* approximation, that is a constant frequency during the measurement time *MT*, only subject to random noise, but without drift or intentional modulation.

A frequency source with a frequency drift, can be described as:

$$f(t) = f_0 + f_d(t) \qquad [6.1]$$

f_0 is the start frequency value

$f_d(t)$ is the frequency drift over time, with mean value ≠ 0 Hz

If $f_d(t)$ has a *linear* drift with time, then $f_d(t) = d \cdot t$, where d is the frequency drift rate (Hz/s). A linear drift during the measurement would result in an accumulated phase $\phi(t)$ vs time relation that is expressed as:

$$\Phi(t) = 2\pi f_0 t + \pi d t^2 + \Phi_0 \qquad [6.2]$$

Figure 9 shows an example with an exaggerated frequency drift, according to [6.2]. This is obviously a 2:nd order function and should ideally be approximated with a 2:nd order polynomial and not a 1:st order straight line.

The main advantage of the linear regression method is to reduce the influence of noise from the measurement process and superimposed random noise on the test signal, and thus increase resolution.

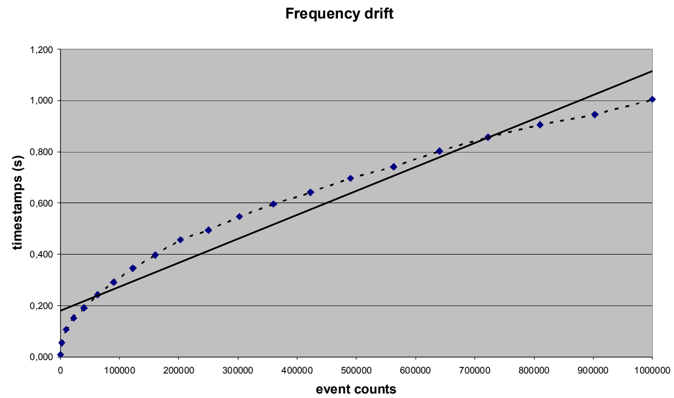

Figure 9. A regression line does not improve resolution if there is a frequency drift

Please note that the linear regression method only can improve the *frequency* resolution which contains several trigger events during the measurement. For single-shot *time interval* measurement it is the basic resolution of the traditional start-stop measurement t_{RES} that sets the limit.

Finally we can conclude that the *continuous time-stamping* is the key to, and a prerequisite for, the regression analysis. This method is not possible in the 2:nd or 3:rd generation counters.

Continuous timestamping has an other advantage, when it comes to characterizing the short-term stability of stable oscillator clocks. The dominating tool for this characterization is the calculation of *Allan Variance* (*AVAR*) and the *Allan Deviation* (*ADEV*), where $ADEV = \sqrt{AVAR}$.

The calculation is performed via a number of back-to-back (zero-dead-time) frequency measurements over a defined measurement time τ. For theoretically correct calculation of AVAR, to avoid dead-time between measurements, the continuous time-stamping is also a prerequisite. Allan Deviation will be discussed in next section.

7. Allan Deviation and continuous time stamping

A stable clock oscillator has a frequency vs time characteristic that can be described as:

$$f(t) = f_N + f_{offs} + f_d(t) + f_r(t) \qquad [7.1]$$

f_N is the nominal frequency

f_{offs} is the initial frequency offset from nominal (calibration uncertainty)

$f_d(t)$ is the frequency drift over time (long term), with mean value ≠ 0 Hz

$f_r(t)$ is the random variation (short term stability), with mean value = 0 Hz

Typical clock oscillators, $f_d(t)$ have a non-linear drift over very long time periods (years), but for shorter periods (days, weeks) we can assume a linear drift with time, that is

$$f_d(t) \cong Df_N t \qquad [7.2]$$

where D is the fractional frequency drift rate and is assumed to be a constant.

All continuous periodic signals can be regarded as a sum of sine wave signals, so let us have a look at the relation between frequency, phase and time in a sine wave signal.

A continuous sine wave with constant amplitude A, can be expressed as $U(t) = A\sin\Phi(t)$, and its momentary or instantaneous frequency $f(t)$ is expressed as:

$$f(t) = \frac{1}{2\pi} \cdot \frac{d\Phi(t)}{dt} \qquad [7.3]$$

Assuming that the frequency drift is linear with a constant drift rate, we can now combine [7.1], [7.2] and [7.3] to express the total phase of a sine wave signal in terms of the various frequency components. The term φ(t) is the *random* phase variation causing short-term frequency instability $f_r(t)$.

$$\Phi(t) = 2\pi(f_N + f_{offs})t + \pi Df_N t^2 + \varphi(t) + \Phi_0 \qquad [7.4]$$

The *fractional* random frequency deviation from the nominal value is commonly denoted $y(t)$, where:

$$y(t) = \frac{f_r(t)}{f_N} \qquad [7.5]$$

The fractional random frequency deviation is:

$$y(t) = \frac{f_r(t)}{f_N} = \frac{1}{2\pi f_N}\frac{d\varphi(t)}{dt} \qquad [7.6]$$

And it is this random variable that is used to characterize the *short-term stability* of oscillators. Characterizing a stable oscillator includes measurement of all frequency components, nominal, offset, drift and short-term stability. However, measurement of frequency offset and drift can be made in a reasonably straightforward way, by applying a normal frequency measurement over sufficiently long measurement time. The random uncertainty of any frequency counter always improves with measurement time.

Measurement of short-term stability is more challenging, because you need to combine high-resolution measurements with short measurement times. The *Allan Deviation* or *Root Allan Variance* is the commonly accepted method for calculation of the short-term clock stability in the time domain. The common measure is Allan Deviation, which is expressed in Hz and not Hz^2 as is the case for AVAR.

$$ADEV = \sqrt{AVAR} = \sigma_y(\tau)$$

We have seen in [7.6] that the random phase variation $\varphi(t)$ causes the random frequency variation $y(t)$. The random phase variation can in a similar way also be expressed as a *random time deviation*, which is useful for the analysis of Allan Deviation:

$$x(t) = \frac{\varphi(t)}{2\pi f_N} = \frac{T_N \cdot \varphi(t)}{2\pi} \qquad [7.7]$$

The random time variation $x(t)$ at times $t = k \cdot T_N$ is a measure of the deviation between actual (noisy) signal relative to the ideal signal ($\bar{x}(t) = T_N \cdot \overline{\Phi}(t)/2\pi$) at the zero-crossings of the signal.

A true instantaneous frequency $f(t_0)$ or instantaneous fractional frequency $y(t_0)$ at time t_0 is not an observable quantity in practice, unlike e.g. the instantaneous phase ϕ(t_0). The measurement of frequency at start time t_k is always performed as an average value over a certain measurement time (τ) in all measurement equipment.

$$\bar{y}_k(\tau) = \frac{1}{\tau}\int_{t_k}^{t_k+\tau} y(t)dt \text{ or}$$

$$\bar{y}_k(\tau) = \frac{\varphi(t_k+\tau)-\varphi(t_k)}{2\pi f_N \tau} = \frac{x(t_k+\tau)-x(t_k)}{\tau} \qquad [7.8]$$

The Allan Variance is defined as:

$$AVAR = \sigma_y^2(\tau) = \frac{1}{2}\left\langle (\bar{y}_{k+1}(\tau) - \bar{y}_k(\tau))^2 \right\rangle \qquad [7.9]$$

The Allan Variance AVAR is an estimate of the variations of a clock frequency over a given measurement time (τ) from one averaging period to the next. An Allan Deviation calculation should in theory be made over an infinite number of samples, each frequency sample being measured back-to-back to the previous, without any dead-time between samples. In practice there are no indefinite measurement periods, and the AVAR for a finite number of samples N can be expressed as:

$$AVAR = \sigma_y^2(\tau) \approx \frac{1}{2(n-1)}\sum_{k=0}^{k=n-1}(\bar{y}_{k+1}(\tau) - \bar{y}_k(\tau))^2 \qquad [7.10]$$

Since short-term instability can be expressed arbitrarily as frequency, phase or time deviation [7.8], let us look at the AVAR expressed as random time deviation instead of random frequency deviation.

$$\sigma_y^2(\tau) \approx \frac{1}{2\tau^2(n-1)}\sum_{k=0}^{k=n-2}(x(t_k+2\tau) - 2x(t_k+\tau) + x(t_k))^2 \qquad [7.11]$$

This way of calculating Allan Variance is suitable for time stamping counters, with pacing period τ, because $x(t_k) = x(t_0 + k \cdot \tau)$ is simply the timestamp of the signal's zero-crossings, sampled every τ seconds, during the total measurement duration *n*τ. A counting technique based on continuous time-stamping will also automatically lead to *zero-dead-time frequency measurements*, which is according to the underlying theory and definition. The CNT-90 Timer/Counter/Analyzer from Pendulum Instruments operates in this mode, enabling correct ADEV measurements and calculation.

Traditional counters, using start-stop frequency measurements over measurement time τ, will always cause a dead-time between measurements. No matter how short this dead-time is, it will be present with at least one cycle, due to the inherent counter design.

Figure 10. The CNT-90 Timer/Counter/Analyzer from Pendulum Instruments AB

References

[Ref 1] Enrico Rubiola, On the measurement of frequency and of its sample variance with high-resolution counters, arXiv physics/0411227 v2 31 Dec2004

[Ref 2] Stefano Bregni, Synchronization of Digital Telecommunications Networks, John Wiley and Sons Ltd 2002, 203-272

[Ref 3] Hewlett-Packard, Application Note 1289, The Science of Timekeeping, Hewlett-Packard 1997

[Ref 4] Pendulum Instruments AB, CNT-90 Timer/Counter/Analyzer User's manual, rev 2 2005

Author Biography

Staffan Johansson, born 1948, achieved his M.Sc. degree in Applied Physics at the Royal Institute of Technology (KTH) in Stockholm, Sweden in 1973. He has been active in the Test & Measurement business since 1981, when he joined the Philips development centre in Stockholm, for counters and pulse generators. Pendulum Instruments AB is a spin-off company from these Philips T&M activities (1998).

Staffan Johansson has made several international presentations and seminars in the field of counter technology, and is now responsible for Marketing, sales and New Product Definitions in the company.